RESEARCH DEPARTMENT

The design of receivers for the pilot-tone stereophonic system

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THE DESIGN OF RECEIVERS FOR THE PILOT-TONE STEREOPHONIC SYSTEM

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SUMMARY

This report discusses the design of receivers for the pilot-tone (Zenith-G.E.) stereophonic broadcasting system with particular reference to the requirements for minimizing susceptibility to interference. The design of the tuner, i.e. that portion of the receiver up to and including the discriminator, is discussed in general terms only. The design of the decoder, i.e. the portion which derives the two separate audio-frequency signals from the multiplex waveform, is considered in more detail. By way of example the details and performance of a particular design of decoder are given.

1. INTRODUCTION

Stereophonic reproduction, in all of the methods currently considered or used for broadcasting or recording for domestic purposes, involves two separate audio-frequency signals which are intended to be reproduced from a left-hand and a right-hand loudspeaker. In the pilot-tone system for stereophonic broadcasting these signals, designated A and B respectively, are transmitted as $M = \tfrac{1}{2}(A+B)$ and $S = \tfrac{1}{2}(A-B)$.

The S signal is first amplitude modulated on to a 38 kc/s sub-carrier, the sub-carrier component being suppressed, and the resulting sideband spectrum is then added to the audio-frequency M signal. In order that the receiver may re-generate the 38 kc/s sub-carrier, which is necessary to demodulate the S signal, a pilot tone at a frequency of 19 kc/s is also transmitted. The sum of these three components, the so-called multiplex signal, comprises the modulation waveform applied to the transmitter.

The relative magnitudes of the components of the multiplex signal and the phase relationship between the pilot tone and the suppressed sub-carrier can best be expressed by the following equation for the instantaneous deviation of the transmitted carrier:

$$\Delta f = 0{\cdot}9\left[\frac{A+B}{2} + \frac{A-B}{2}\sin\omega_s t + 0{\cdot}1\sin\frac{\omega_s}{2}t\right] 75 \text{ kc/s}$$

$$= 0{\cdot}9\left[M + S\sin\omega_s t + 0{\cdot}1\frac{\omega_s}{2}t\right] 75 \text{ kc/s}$$

where A and B vary within the range ± 1 and $\omega_s/2\pi$ = 38 kc/s. High-frequency pre-emphasis (50 μs time constant) is employed, as in monophonic f.m. broadcasting, and the symbols A, B, M and S here denote pre-emphasized signals.

The spectrum of the multiplex signal is shown in Fig. 1.

Conventional monophonic v.h.f. sound receivers will reproduce only the M signal. It is generally agreed that this gives an entirely acceptable version of the programme from a single loudspeaker and that the presence of the ultrasonic components in the modulation waveform does not impair the operation of the receiver.

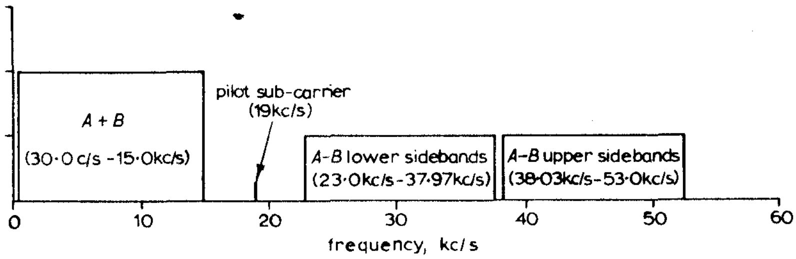

Fig. 1
Spectrum of pilot-tone system multiplex waveform

The only effect on monophonic reception which would result from changing the v.h.f. broadcasting system from monophony to the pilot-tone system would be a slight reduction (about 2 dB)[1] in signal-to-noise ratio consequent upon the reduction of the transmitter deviation by the M signal.

A stereophonic receiver is required to carry out an additional demodulation process in order to obtain the S signal and, in so doing, it will also demodulate those noise components in the discriminator output which lie in the region of the sub-carrier frequency and which, in a monophonic receiver, would produce no audible output. As a result, the signal-to-noise ratio of a stereophonic system employing an ultrasonic sub-carrier modulated by a programme signal is lower than that of a monophonic system[2] and this factor tends to reduce somewhat the service area of a transmitter for stereophonic reception. In order to provide the best possible service it is therefore important that stereophonic receivers should be designed not only to give satisfactory reproduction with an immaculate input signal but also to prevent any *avoidable* degradation of performance when the input signal is disturbed by noise, interference or propagation effects.

2. THE RECEIVER

2.1. General Requirements for Stereophonic Reception

It is convenient to consider a receiver for the pilot-tone system in two sections; the tuner, i.e., that part up to and including the discriminator, which selects, amplifies and demodulates the incoming frequency-modulated signal to produce the multiplex waveform, and the subsequent circuits which are concerned with decoding the multiplex signal to recover the separate A and B audio-frequency outputs.

The vector diagrams in Fig. 2 show the signals at various stages through the system with two separate conditions of modulation. From this it can be seen that, for correct reconstitution of the A and B signals or 'de-matrixing', the relative phases and amplitudes of the sum and difference signals must be preserved.

In Fig. 2(b), for example, which represents a modulation input applied to the A channel only, if the difference vector $\frac{1}{2}(A-B)$ were shifted in phase or changed in

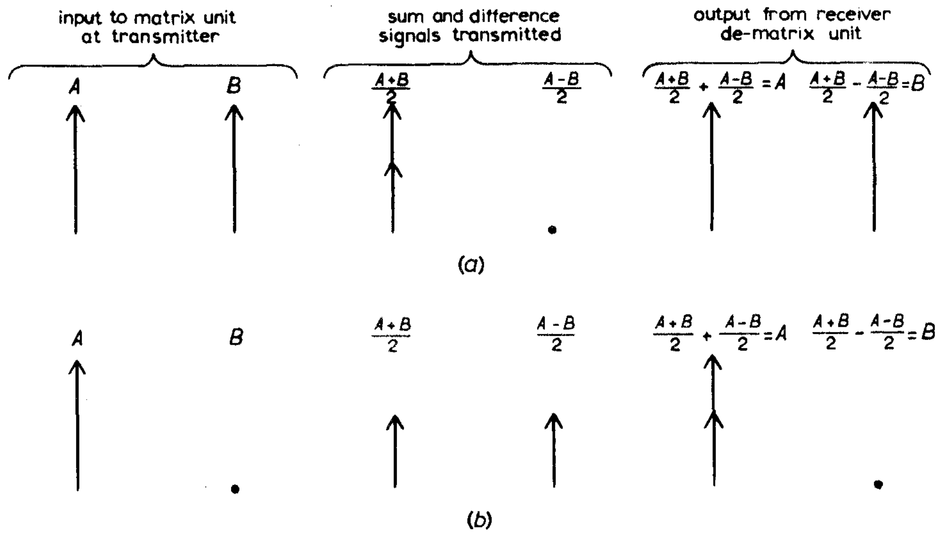

Fig. 2 - Vector relationships of signals in an ideal pilot-tone stereophonic system

(a) Inputs to A and B channels equal and in phase
(b) Input to A channel only

amplitude relative to the sum vector $\frac{1}{2}(A+B)$, the resultant B signal, ideally $\frac{1}{2}(A+B)$ - $\frac{1}{2}(A-B)$, would not be zero and there would be cross-talk from the A channel into the B channel output.

In designing the tuner for a stereophonic receiver, two characteristics of the pilot-tone system should be borne in mind:

(i) the increase in the transmitted modulation-frequency bandwidth from 15 kc/s to 53 kc/s,

(ii) the increased susceptibility to noise and interference.

Ideally, the overall response at the discriminator output should be uniform in amplitude and linear in phase up to 53 kc/s. Any distortion of the multiplex spectrum occurring here will have precisely the same effect as the distortion caused by the decoder low-pass filter discussed in Section 2.2.1, and the resulting cross-talk can be evaluated by the formulae derived in the Appendix. One factor affecting the overall response is the amplitude characteristic of the intermediate frequency amplifier. It is not always appreciated that at the higher frequencies for which the modulation index is of the order of unity or less the modulation-frequency response of a f.m. receiver depends upon the i.f. response over the bandwidth occupied by the first-order modulation sidebands in a manner very similar to that of an a.m. receiver.[3] Attenuation of the higher frequency components of the multiplex waveform due to this cause could be removed by increasing the i.f. bandwidth but this expedient is to be regarded with caution since it increases the possibility of adjacent-channel interference, an aspect of performance in which the pilot-tone system is in any case only marginally acceptable.

Generally speaking, quite large departures from the ideal modulation-frequency response at the discriminator output can be tolerated since the mean difference in level between the sum and difference-sideband signals can be compensated, as discussed in Section 2.2.1.

In view of the reduction in signal-to-random-noise ratio with this system, the tuner noise factor is important in determining the minimum signal level at which satisfactory reception is possible. Owing to the presence of cosmic noise the effective aerial temperature in Band II is some 1200° - 1500°K and it is therefore not possible to realize the full benefit of a very low receiver noise factor.[4] For example, a true receiver noise factor of 6 dB would give an effective noise factor of between 8·5 dB and 9 dB when allowance for cosmic noise is made.

Efficient amplitude limiting is essential in a stereophonic tuner in order to remove the avoidable amplitude modulation component of any interference which may be present. It is also necessary to ensure that the tuner output level remains within the range in which satisfactory operation of the decoder is obtained for all usable r.f. signal levels. This presents no great difficulty; r.f. signals below one or two hundred microvolts would generally give an unacceptable signal-to-noise ratio, and with inputs above this level the receiver a.g.c. is normally fully operative.

It may be said that, in general, the requirements of the tuner are similar in type of those applicable for monophonic receivers although more stringent in degree. The novel part of the receiver is that concerned with decoding the multiplex signal.

2.2. The Decoder

The functions of the decoder are:

(i) To regenerate from the 19 kc/s pilot tone the 38 kc/s sub-carrier which is suppressed in transmission.

(ii) To demodulate the difference signal.

(iii) To dematrix the sum and difference signals, i.e. re-combine them to recover the original *A* and *B* channel modulation signals.

Noise and interference accompanying the radio-frequency input signal to the receiver will produce unwanted components in the multiplex input applied to the decoder. As far as components lying within the frequency bands occupied by the sum signal and by the difference-sideband signal are concerned, they will be demodulated in precisely the same way as the wanted signal to produce an audible output; the design of the decoder can have no effect on this. Decoder design does, however, affect the response of the receiver to those components of interference and noise which occur above 53 kc/s and in the region of the pilot-tone frequency. Unless appropriate precautions are taken, frequencies above the limit of the normal multiplex spectrum can intermodulate with harmonics of the sub-carrier to produce an audio-frequency output in the difference channel. Similarly, components of interference close to 19 kc/s can enter the sub-carrier regenerating circuits and produce unwanted modulation of the sub-carrier; this also can give rise to an audio-frequency output.

Although the sub-carrier regenerator might appear to be the logical starting point in a discussion of decoders it will be more convenient to deal first with the basic operation of demodulating the multiplex signal.

2.2.1. The Switching Decoder

It is possible to combine operations (ii) and (iii) and produce the *A* and *B* signals directly from the multiplex signal without a separate dematrixing process. The type of decoder which operates on this principle is somewhat simpler than that in which the sum and difference signals are extracted separately, and will be described first.

Fig. 3 shows the decoder in block schematic form. The multiplex input is fed to two demodulators, together with demodulating signals $f_1 f_2$ at the sub-carrier

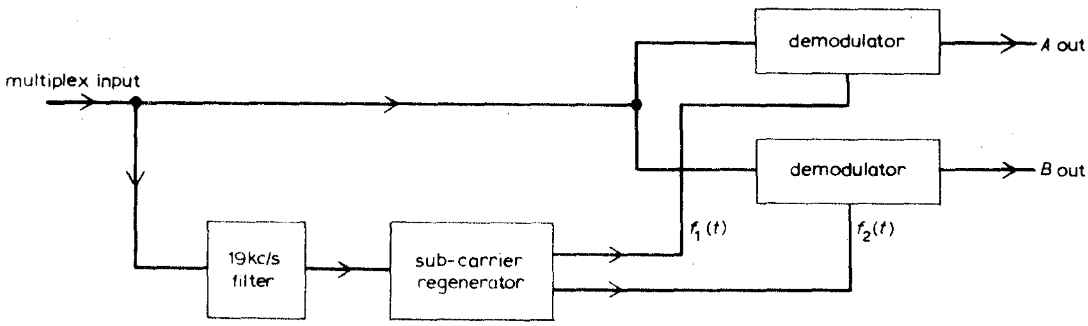

Fig. 3 - Switching decoder, block schematic

frequency, and the *A* and *B* outputs are produced directly. The general expression for the multiplex waveform as given in Section 1, is:

$$\frac{A+B}{2} + \frac{A-B}{2} \sin \omega_s t + 0{\cdot}1 \sin \frac{\omega_s t}{2}$$

where *A* and *B* represent the audio-frequency input signals to the left- and right-hand channels respectively and $\omega_s/2\pi$ is the sub-carrier frequency, 38 kc/s.

If the demodulators were linear multiplying devices and

$$f_1(t) = \tfrac{1}{2} + \sin \omega_s t$$

$$f_2(t) = \tfrac{1}{2} - \sin \omega_s t$$

then the outputs of the demodulators would be

$$E_A = \left[\frac{A+B}{2} + \frac{A-B}{2} \sin \omega_s t + 0{\cdot}1 \sin \frac{\omega_s t}{2}\right] \left[\tfrac{1}{2} + \sin \omega_s t\right]$$

$$E_B = \left[\frac{A+B}{2} + \frac{A-B}{2} \sin \omega_s t + 0{\cdot}1 \sin \frac{\omega_s t}{2}\right] \left[\tfrac{1}{2} - \sin \omega_s t\right]$$

Considering only the audio-frequency terms in the products

$$E_A = \frac{A+B}{4} + \frac{A-B}{4} = \frac{A}{2}$$

$$E_B = \frac{A+B}{4} - \frac{A-B}{4} = \frac{B}{2}$$

The coefficients of the terms representing the contributions of the sum and difference signals in each channel are numerically equal. Perfect stereophonic demodulation is thus achieved with the assumed form of $f_1(t)$ and $f_2(t)$, as has been pointed out by De Vries.[5]

This multiplication process is a precise operation which, by domestic receiver standards, would probably be expensive to realize in practice. An attractive alternative, however, is to use two simple diode gating circuits operating as on-off switches with a 1/1 mark-to-space ratio. This is equivalent to multiplication by a square wave, for which case the output signals are:[5]

$$E_A = \left[\frac{A+B}{2} + \frac{A-B}{2}\sin\omega_s t + 0{\cdot}1\sin\frac{\omega_s t}{2}\right]\left[\tfrac{1}{2} + \frac{2}{\pi}\sin\omega_s t + \frac{2}{3\pi}\sin 3\,\omega_s t\ \ldots\right]$$

$$E_B = \left[\frac{A+B}{2} + \frac{A-B}{2}\sin\omega_s t + 0{\cdot}1\sin\frac{\omega_s t}{2}\right]\left[\tfrac{1}{2} - \frac{2}{\pi}\sin\omega_s t - \frac{2}{3\pi}\sin 3\,\omega_s t\ \ldots\right]$$

Again considering only the audio-frequency terms in the products,

$$E_A = \frac{A+B}{4} + \frac{A-B}{2\pi} = \frac{\pi+2}{4\pi}A + \frac{\pi-2}{4\pi}B$$

$$E_B = \frac{A+B}{4} - \frac{A-B}{2\pi} = \frac{\pi+2}{4\pi}B + \frac{\pi-2}{4\pi}A$$

Here, the coefficients of the sum and difference terms are not equal, with the result that part of the A channel input is reproduced from the B channel output and vice versa. This is a fundamental result of applying the square-wave switching to the signal, but the cross-talk can be compensated by passing the multiplex signal through a network which attenuates the sum signal by a factor of $2/\pi$ relative to the difference-sideband signal. Alternatively, this compensation can be applied after demodulation, and one method of achieving this is by the use of an audio-frequency amplifier with partial common-mode suppression as shown in Fig. 4. With inputs to the two grids in-phase, as would arise from the sum signal, R_3 has no effect on the gain of the stage.

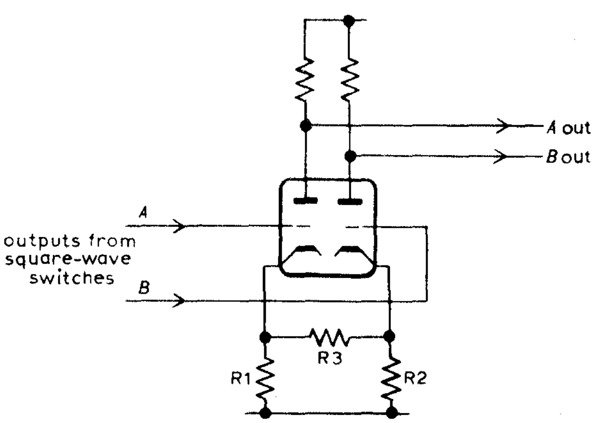

Fig. 4 - Audio-frequency amplifier with partial common-mode suppression

With anti-phase inputs, however, the negative feedback is reduced by the presence of R_3. Thus the gain is greater for the difference-signal component than for the sum-signal component and this gain differential can be adjusted within quite wide limits by suitable proportioning R_1, R_2 and R_3.

The application of this type of circuit is, of course, not limited to correcting the deficiencies of the square-wave switching system. If R_3 is made variable it forms a convenient control to compensate for unbalance between the sum and difference signal amplitudes arising elsewhere in the receiver.

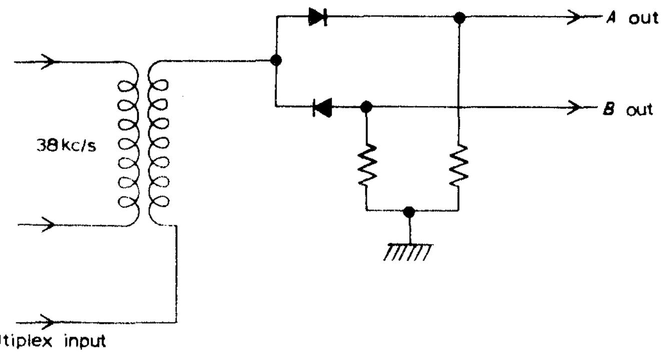

Fig. 5 - Unbalanced switching demodulator

The simplest form of switching demodulator is shown in Fig. 5. The waveform of the switching sub-carrier input is normally sinusoidal but, provided that it is sufficiently large compared with the multiplex input, the operation of the circuit will approximate closely to the square-wave switching condition. This results, however, in a large sub-carrier-frequency component appearing in the output, which, even after de-emphasis, can be an embarrassment in the following a.f. stages. Even more important, any audio-frequency modulation of the regenerated sub-carrier will also be reproduced. Both of these effects can be minimized by the use of

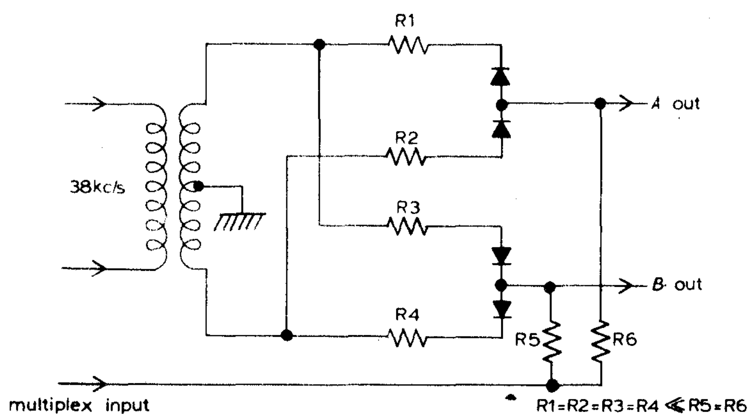

Fig. 6 - Balanced switching demodulator

a balanced switching circuit, such as that shown in Fig. 6, in which the output at the sub-carrier frequency is suppressed to an extent limited only by the degree of balance achieved in the diode bridges. This form of demodulator is also much less susceptible to amplitude modulation of the sub-carrier switching input. Nevertheless, if the sub-carrier input waveform is sinusoidal, the switching waveform is in fact not a true square wave but a clipped sine wave and amplitude modulation of the sub-carrier can produce pulse width modulation of the switching waveform and hence can still have some effect on the demodulated audio output.

In view of the degradation of signal-to-noise ratio which can be caused by interference components above 53 kc/s in the multiplex waveform, it is highly desirable that the demodulators be preceded by a low-pass filter. If this filter is not to introduce differential phase or amplitude errors in the demodulated sum and difference signals, which would result in cross-talk between the *A* and *B* channels, certain requirements must be satisfied. If $r(f)$ is the amplitude response and $\phi(f)$ is the phase response at a frequency f, while f_a and f_s are respectively the audio modulation frequency and sub-carrier frequency, the requirements may be stated as follows:

(i) The phase characteristic must be such that $d/df\ \phi(f)$ has the same value at $f = f_a$, $f = f_s - f_a$ and $f = f_s + f_a$ for all values of f_a.

(ii) The amplitude characteristic must be such that the ratio between $r(f_a)$ and $\frac{1}{2}\left[r(f_s - f_a) + r(f_s + f_a)\right]$ is constant for all values of f_a.

Note: This ratio need not necessarily be unity since any constant inequality between the sum and difference signals can be compensated in the same way as the difference in detection efficiency of the square-wave switch demodulator.

If the filter has constant attenuation and a linear phase characteristic between 30 c/s and 53 kc/s, these requirements are, of course, fulfilled. A practical simple filter with a cut-off frequency in the region of 55 kc/s would approximate quite closely to the required characteristic except in the immediate vicinity of the cut-off frequency. This would produce a phase shift, and possibly some attenuation, of the upper sideband of the difference-sideband signal at high modulation frequencies. Formulae are derived in the Appendix for calculating the resulting phase shift and attenuation of the demodulated difference signal and also the cross-talk which this would produce.

Fig. 7 gives the level of cross-talk, as a function of frequency, which is given in CCIR Report No. 293 (Geneva 1963) as just perceptible in a stereophonic broadcasting system. It shows that the permissible cross-talk increases at

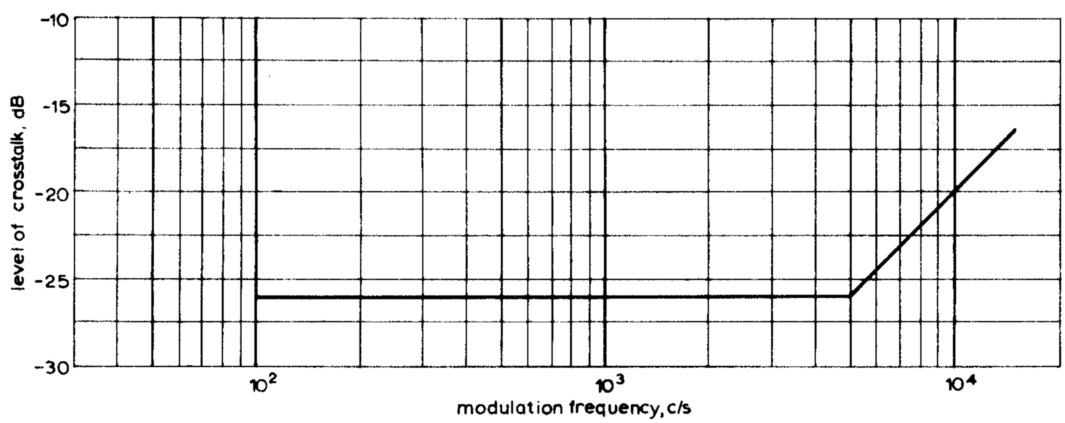

Fig. 7 - Minimum perceptible level of cross-talk between A and B channels

6 dB per octave at frequencies above 5 kc/s, hence there is considerable tolerance of departures from the ideal characteristic in the decoder multiplex signal filter.

2.2.2. The Sub-Carrier Regenerator

The requirement to be satisfied by the sub-carrier regenerator is that the 38 kc/s output should be in the correct phase to produce maximum output from the demodulators, free from both long-term and audio-frequency phase variations and from audio-frequency amplitude modulation. The output of the difference signal component from the demodulators is proportional to $\cos\psi$, where ψ is the phase error of the switching sub-carrier (see Appendix). The output of the sum signal component is, however, independent of ψ. Long-term phase drift will thus produce cross-talk between the *A* and *B* channels, and phase modulation at an audible frequency will produce amplitude modulation of the difference signal.

In view of the fact that the difference signal amplitude follows a cosine law against the sub-carrier phase error, it is important that the initial phase error is small in order to minimize the disturbance to the output arising from any subsequent drift or phase modulation. It is therefore undesirable to use variation of sub-carrier phase as a means of making adjustments to the difference-signal amplitude.

To minimize disturbances of the 38 kc/s output by interference, some means of amplitude limiting should be incorporated in the sub-carrier regenerating channel and the bandwidth of the filter circuits should be as small as is practicable. This second parameter is a matter for compromise since the requirement for minimum susceptibility to interference conflicts with that for good long-term phase stability.

Either of two methods can be adopted for regenerating the 38 kc/s sub-carrier. The pilot-tone can be used to lock a local oscillator at the pilot or sub-carrier frequency or it can be applied to some form of frequency doubler. In principle, it is probably easier to obtain good amplitude-modulation suppression with the locked-oscillator but this type of circuit is more prone to produce phase modulation of the output as a result of amplitude modulation of the locking signal. It is difficult to generalize, since the performance is determined more by the details of design than by the type of circuit adopted.

2.2.3. The Sum-and-Difference Separation Decoder

Fig. 8 shows the block schematic diagram of the type of decoder in which demodulation of the difference signal and de-matrixing are carried out separately. The difference-sideband signal is extracted from the multiplex input by the high-pass filter, added to the regenerated 38 kc/s sub-carrier, and the *A-B* audio-frequency signal obtained by rectification in a mean or envelope detector. The delay network in the *A+B* channel is required to compensate for the delay introduced by the high-pass filter. In the de-matrix network the sum and difference signals are combined to form the *A* and *B* outputs.

This form of decoder embodies some additional circuit features as compared with the switching type. It requires additional filtering to separate the sum and difference signals, and the resultant differential delay in the sum and difference channels has to be corrected. If a simple diode detector is used a higher degree of amplitude modulation suppression is required in the sub-carrier regenerating channel, since the difference detector will treat a.m. sidebands accompanying the sub-carrier in precisely the same way as components of the wanted difference-signal modulation.

As in the switching decoder, a low-pass filter with a cut-off frequency of about 55 kc/s is highly desirable to reduce the effects of interference. The effects of phase shift of the sub-carrier and distortion of the spectrum of the difference-sideband signal are sensibly identical with both types of decoder circuit, provided that the re-introduced sub-carrier in the sum-and-difference decoder is

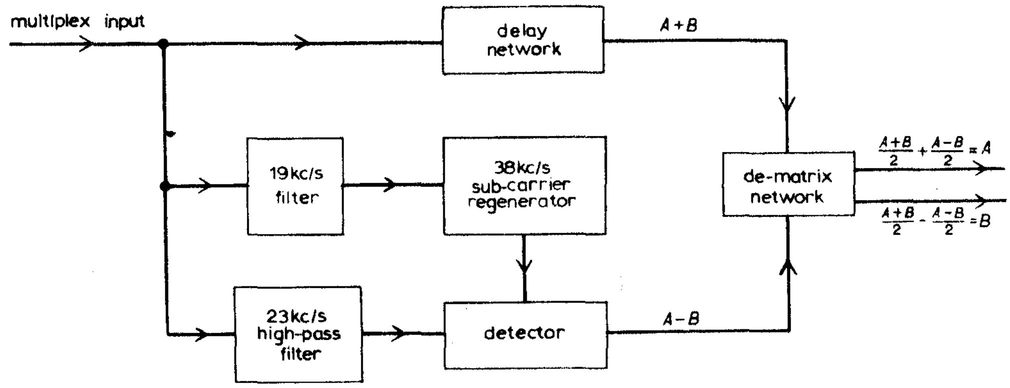

Fig. 8 - Sum-and-difference separation decoder

large relative to the peak difference-sideband signal. If the sub-carrier amplitude is less than two or three times that of the peak double-sideband signal the forms of distortion which are associated with asymmetrical sideband reception of a.m. may become appreciable.

2.3. A Practical Decoder Design

Having discussed the general requirements of decoders, the details can best be illustrated with reference to a specific design. Fig. 9 shows the circuit diagram of a switching decoder.

The 19 kc/s pilot tone is extracted from the multiplex signal in the cathode circuit of V_{1A} by the single-circuit filter L_1, C_3. V_2 is a linear amplifier, the anode load being a second 19 kc/s tuned circuit L_2, C_9. Diodes MR1 and MR2 form the frequency doubler and V_{1B} functions as a saturated grid limiter and 38·kc/s amplifier. The resonant primary circuit of the transformer T_1 gives further filtering of the regenerated sub-carrier and a sinusoidal switching voltage of 40V peak-to-peak is developed across the secondary of T_1. Fig. 10 shows the static limiting characteristic of the sub-carrier regenerating channel.

The multiplex signal is taken from the anode of V_{1A} through the low-pass filter C_5, L_3, C_6 and the feed resistors R_{15} and R_{16} to the switching bridges. The 19 kc/s tuned circuit L_1, C_3 in the cathode circuit of V_{1A} also functions as a notch filter and reduces the amplitude of the pilot-tone component in the multiplex waveform fed to the switching circuits. This, in turn, reduces the amplitude of any unwanted beat tones arising from intermodulation between the pilot tone and the modulation components of the multiplex signal due to non-linearities in the demodulators or subsequent stages. The measured overall amplitude and phase characteristics of the multiplex signal channel are shown in Figs. 11 and 12.

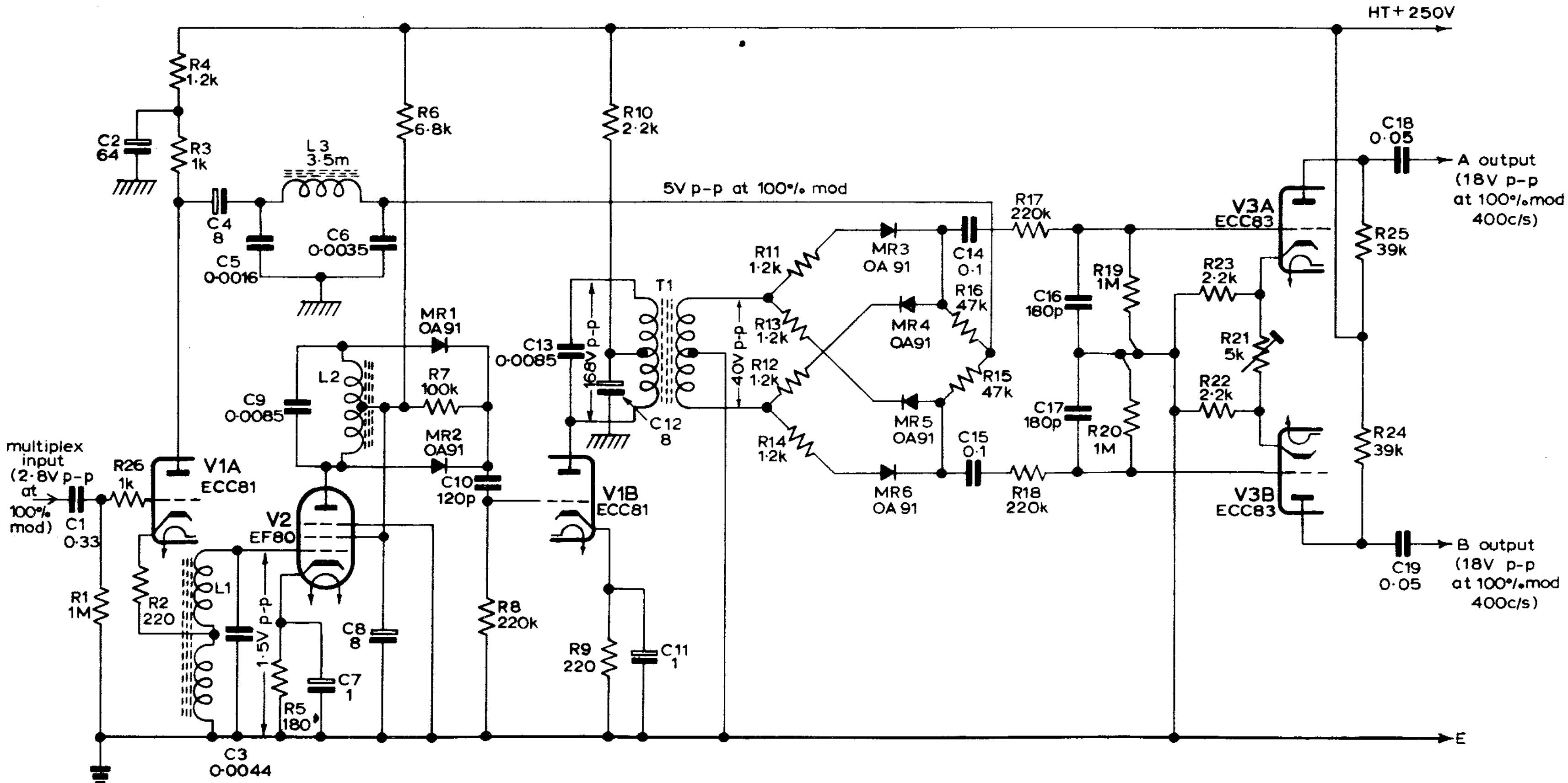

Fig. 9 - Switching decoder, circuit diagram

Considering the phase response, a linear phase characteristic merely represents a constant time delay to all components of the signal and hence has no effect on the process of stereophonic demodulation. In evaluating the cross-talk

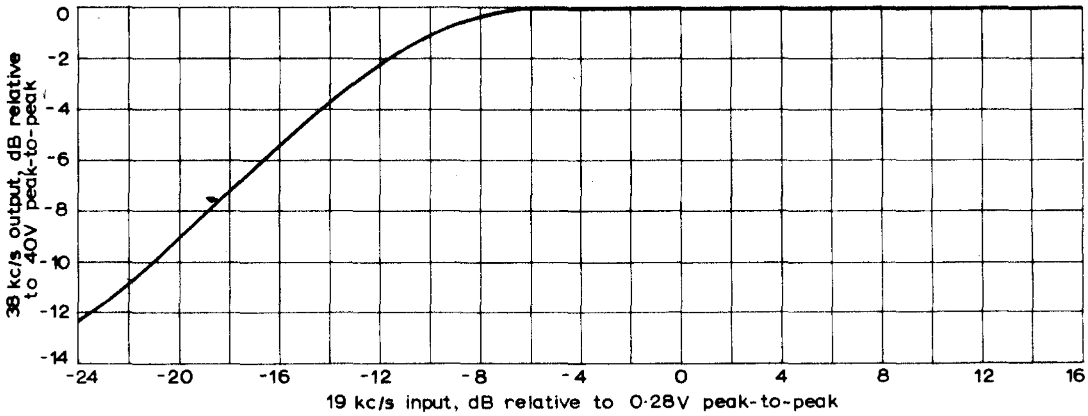

Fig. 10 - Static limiting characteristic of sub-carrier regenerating channel

which will result from imperfections in the response of the multiplex channel, the phase errors to be taken into account are those which represent departures from the linear phase condition.

In this case the maximum errors occur at the highest modulation frequency 15 kc/s. Putting the appropriate values from Figs. 11 and 12 into the expression given in the Appendix, the phase shift and attenuation of the demodulated difference signal may be determined; we have:

phase error at 23 kc/s, $\theta_1 = 14°$

phase error at 53 kc/s, $\theta_2 = -7°$

amplitude factor at 23 kc/s,* $k_1 = 0{\cdot}99$

amplitude factor at 53 kc/s,* $k_2 = 0{\cdot}97$

From these values we find that the demodulated difference signal at 15 kc/s will be reduced by 0·2 dB and phase-retarded by 10·6°.

However, we see from Figs. 11 and 12 that, at 15 kc/s, the sum signal is reduced by 0·4 dB and retarded by 9°. Thus the differential amplitude and phase error is 0·2 dB and 1·6° which, if it were the only imperfection in the circuit, would produce a cross-talk level of -36 dB.

The source impedance presented to the switching circuits by the transformer T_1 is 1250Ω; the series resistors R_{11} to R_{14} prevent an undue drop in the secondary voltage due to loading by the short-circuited pair of diodes. The presence of these resistors, together with the finite resistance of the switching diodes in the nominally short-circuited condition, also produces slight in-phase cross-talk between the *A* and *B* channels but this can be corrected in the following stage V_3.

* The constant attenuation of 0·2 dB over the difference-sideband signal portion of the spectrum has been ignored since it can be compensated in the audio-frequency amplifying stage.

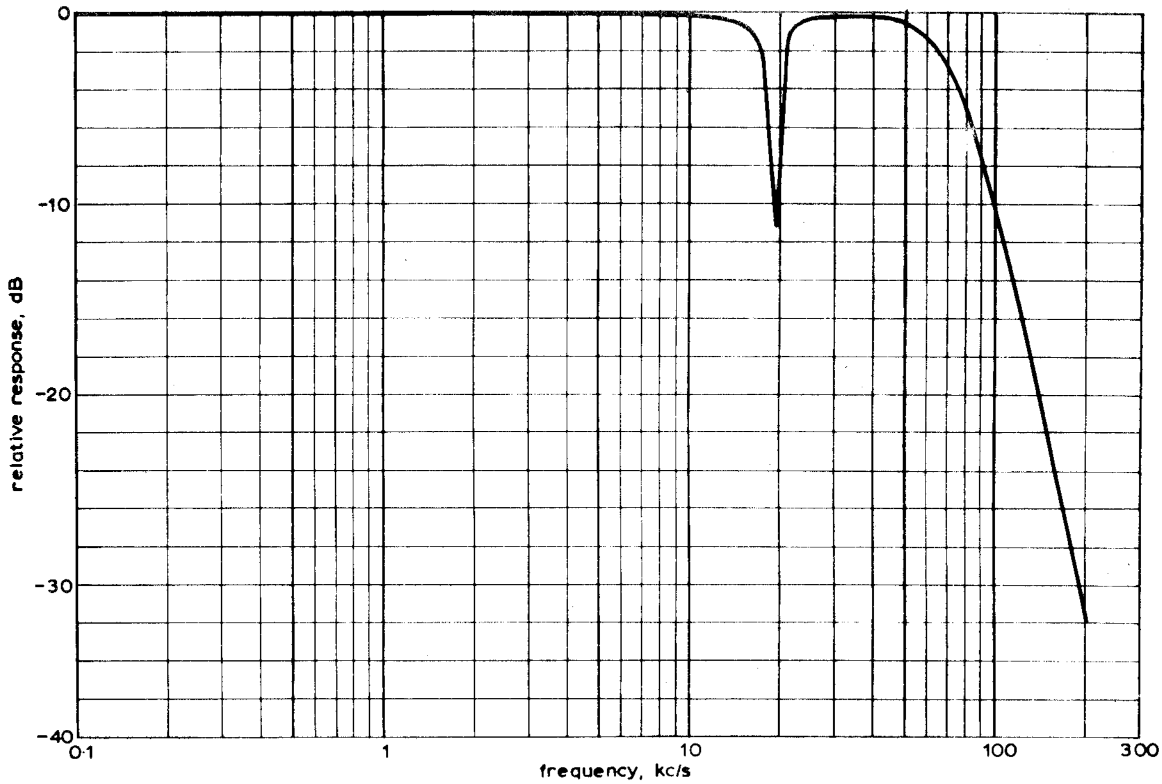

Fig. 11 - Amplitude response of multiplex signal channel

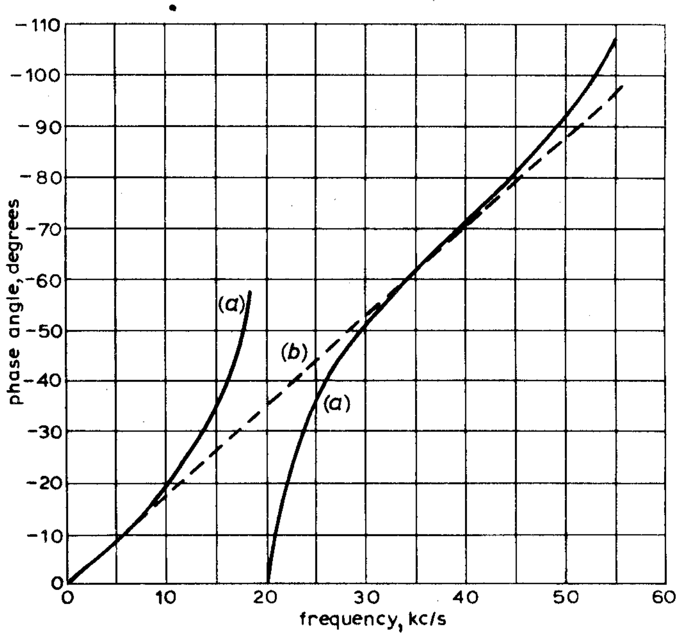

Fig. 12 - Phase characteristic of multiplex signal channel

(a) Measured characteristic
(b) Ideal linear phase characteristic

The de-emphasis circuits precede the first audio-frequency amplifier V_3 and reduce the amplitude of the ultrasonic components in the input to this stage. Precise matching of the time constants is required in order to preserve a constant ratio of common-mode to differential-mode gain over the entire modulation-frequency range.

2.4. Performance

The performance figures given are for the decoder when fed with a 2·8V peak-to-peak multiplex signal from a stereophonic coder. If the decoder were embodied in a complete receiver some deterioration in overall performance might well result from imperfections in the tuner.

Figs. 13 to 16 and Table 3 show the performance test results. The modulation-frequency response curve shown in Fig. 13 is the average of both *A* and *B* channels, since the maximum difference between them was 0·2 dB. The cross-talk shown in Fig. 14 is rather worse at the extreme low and high modulation frequencies than can

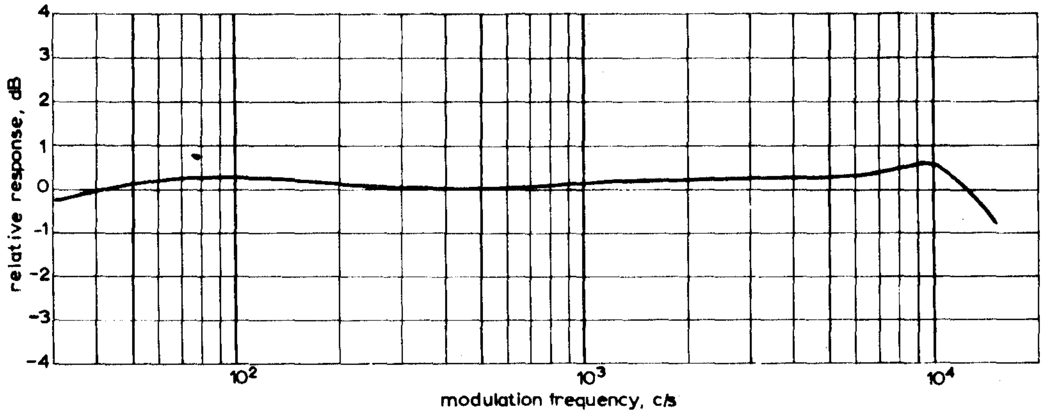

Fig. 13 - Modulation-frequency response of A and B outputs

be accounted for by the characteristics of the multiplex channel alone. This result represents the sum of various instrumentalities in both coder and decoder but, since it is well within the subjective tolerances shown in Fig. 7, it was not investigated further.

TABLE 3

Harmonic distortion with 400 c/s modulation

PERCENTAGE MOD. DEPTH	CHANNEL TO WHICH MODULATION APPLIED	PERCENTAGE DISTORTION AT OUTPUT OF CHANNEL	
		A	*B*
100	*A*	0·7	-
"	*B*	-	0·8
"	*M*	1·6	1·5
"	*S*	0·7	1·0
40	*A*	0·7	-
"	*B*	-	0·7
"	*M*	0·7	0·7
"	*S*	0·6	0·6

Susceptibility to interference components in the multiplex signal in the region of 19 kc/s is shown in Fig. 15, and the method of measurement requires some explanation. Any interference sidebands accompanying the regenerated sub-carrier can produce a disturbance of the audio-frequency output of the receiver either by direct demodulation (as, for example, with an amplitude modulated sub-carrier in a sum-and-difference separation decoder) or by intermodulation with the wanted signal (as, for example, with a phase modulated sub-carrier). In order to reveal the presence of both mechanisms a test signal having 40% modulation at 640 c/s in the

A channel alone was applied together with an interfering sinusoidal signal at a

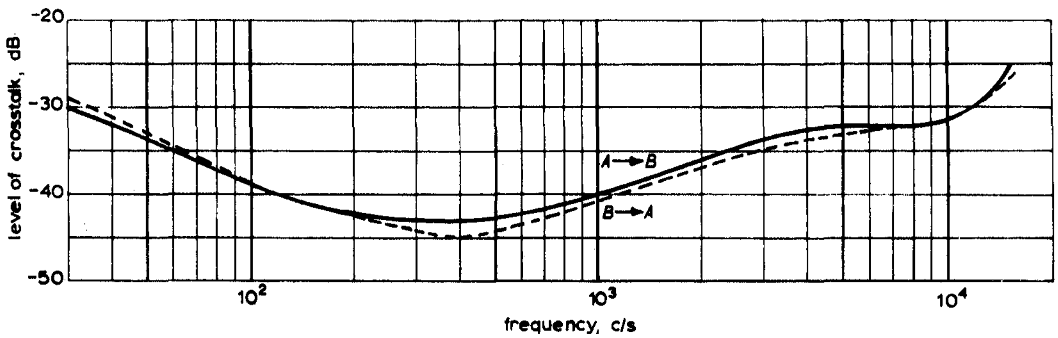

Fig. 14 - Cross-talk characteristic between A and B channels

frequency f_I, f_I being varied in steps from 16 kc/s to 22 kc/s. At each step the level of interference was increased from zero until the largest unwanted component of the receiver output (ignoring components at harmonics of the wanted modulation frequency) was -40 dB relative to the output level of the wanted 640 c/s modulation. The level of the interfering signal f_I was then recorded. Fig. 15 shows the results of the test with, for comparison, results of a similar test on a switching decoder with little or no limiting in the sub-carrier channel and unbalanced single diode switching circuits.

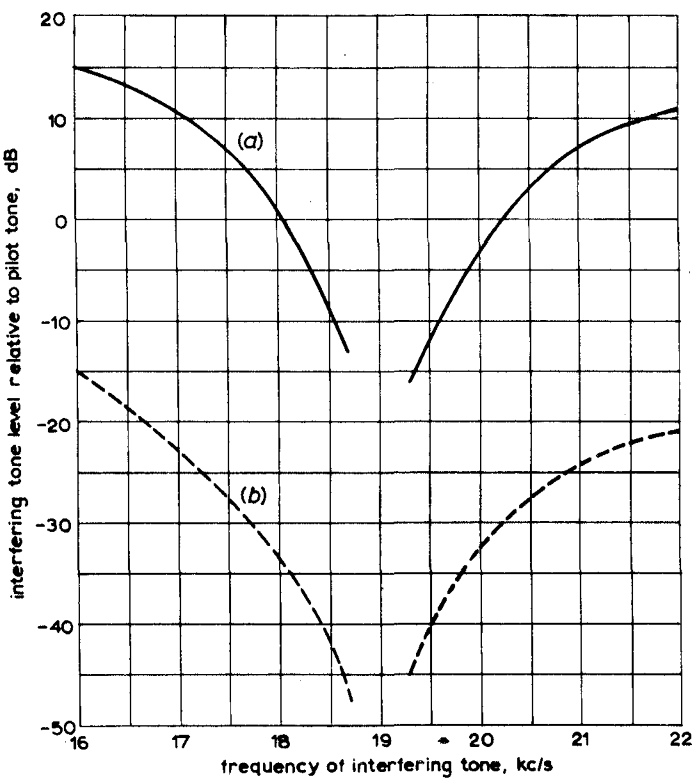

Fig. 15 - Susceptibility to interference in sub-carrier regenerating channel

(a) With adequate sub-carrier limiting
(b) With no sub-carrier limiting

Susceptibility to unwanted components of the multiplex signal at frequencies above 53 kc/s was tested by measuring the 1 kc/s output obtained with an input signal of constant amplitude set, in turn, at frequencies of f_s + 1 kc/s, $2f_s$ + 1 kc/s....nf_s + 1 kc/s, where f_s is the sub-carrier frequency (38 kc/s). The results are shown in Fig. 16 with, for comparison, results of a similar test on the same decoder but with the low-pass filter C_5, L_3, C_6 removed.

The results for total harmonic distortion are given in Table 3.

The performance is regarded as generally satisfactory. One way in which it could probably be improved is by redesigning the multiplex channel low-pass filter

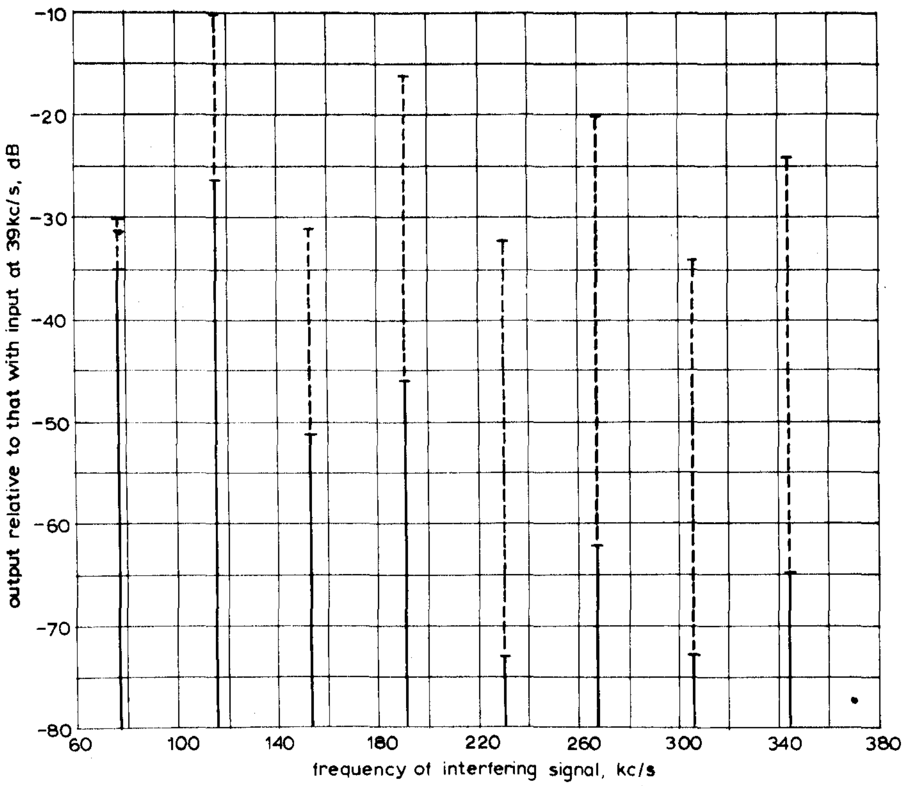

Fig. 16 - Susceptibility to interference from multiplex signal components above 53 kc/s

——— With low-pass filter ----- Without low-pass filter

with a somewhat lower cut-off frequency. This would give increased rejection of interference in the region of 70 to 150 kc/s, at the expense of some increase in cross-talk at the higher modulation frequencies which could well be tolerated.

3. INTERFERENCE

Frequent references have been made in the foregoing discussion to the importance of those aspects of stereophonic receiver design which affect susceptibility to interference and it is of interest to consider some types of interference in more detail.

3.1. Adjacent-Channel Interference

The presence of an interfering transmission spaced in frequency by Δf from that to which the receiver is tuned will give rise to a spectrum of interference, in the output of the discriminator, centred on the frequency Δf and extending outwards to

an extent determined by the modulation of the two transmissions. Where $\Delta f = 200$ kc/s, almost all of this interference spectrum lies outside the bandwidth occupied by the wanted multiplex signal. Nevertheless, an audible output will be produced if any components of the interference are demodulated by beating with harmonics of the sub-carrier frequency. In such a case, the provision of a low-pass filter in the multiplex signal channel would be expected to give a very substantial improvement. Where $\Delta f = 100$ kc/s the improvement, though significant, would be smaller since the interference spectrum will now overlap that of the wanted multiplex signal and cannot be entirely eliminated without disturbance of the wanted modulation. Subjective tests with a very simple switching decoder, fed from the discriminator of a typical medium-priced domestic f.m. receiver, showed an improvement of 15 dB where $\Delta f = 200$ kc/s and 7 dB where $\Delta f = 100$ kc/s by the use of a filter with a response above 40 kc/s similar to that shown in Fig. 11. The improvement quoted represents the increase in the level of the interfering signal required to produce an output interference subjectively rated as 'perceptible'.

3.2. Multipath Propagation

The distortion arising from multipath propagation,[6] that is when the received signal consists of a number of components which have travelled by paths of different lengths and hence suffered different time delays, has been found rather more severe with the pilot-tone stereophonic system than with the normal monophonic system. This is particularly so where the reflected signals involved have time delays corresponding to comparatively small path differences, 5 miles (8 km) or less.

One effect of the presence of signals with path differences close to 5 miles (8 km) is to produce a form of intermodulation in which the pilot tone is amplitude modulated by the low-frequency components of the sum signal. If this modulation is not adequately suppressed by amplitude limiting in the sub-carrier regenerating channel it produces distortion of a particularly unpleasant character. Objective measurements have been made to determine the extent of the improvement which sub-carrier amplitude limiting can provide and the results of one test are given in Fig. 17. They show the amplitude of harmonics produced by two decoders when fed from the discriminator output of a medium-priced f.m. receiver. The r.f. signal input to the receiver contained one delayed component with an amplitude of 10%, and a delay corresponding to a path difference of 5 miles (8 km), relative to the main signal; the signal was modulated 100% at a frequency of 120 c/s in the *A* channel alone. Both decoders were switching types, one with a high degree of limiting, and the other with little or no limiting, in the sub-carrier regenerating channel. (Fig. 15 gave the results of tests on the same two decoders). It can be seen that, although some distortion remains even with the better decoder (this due largely to unavoidable distortion of the sum- and difference-signal information), the total distortion has been very substantially reduced. Two points must be emphasized with regard to the results given in Fig. 17. The first is that the condition measured, with a low modulation frequency and a single delayed signal having a delay corresponding to a path difference of 5 miles (8 km), is that for which the difference between adequate and inadequate sub-carrier limiting is most marked. The second is that it presents a somewhat over-pessimistic picture of the total effect of multipath propagation. The amplitudes of individual harmonics vary very rapidly with small changes in the relative phase of the reflected and direct signals. For the purpose of the test, this phase was adjusted to produce the maximum amplitude of each harmonic in turn and that maximum value is shown in the diagram. In a practical situation, therefore, the harmonics will not all reach the indicated values simultaneously.

3.3. Impulsive Interference

This form of interference might be quite serious to listeners to a stereophonic broadcasting service at the fringe of the transmitter service area. Subjective tests indicate that, with a high level of interference, the difference between

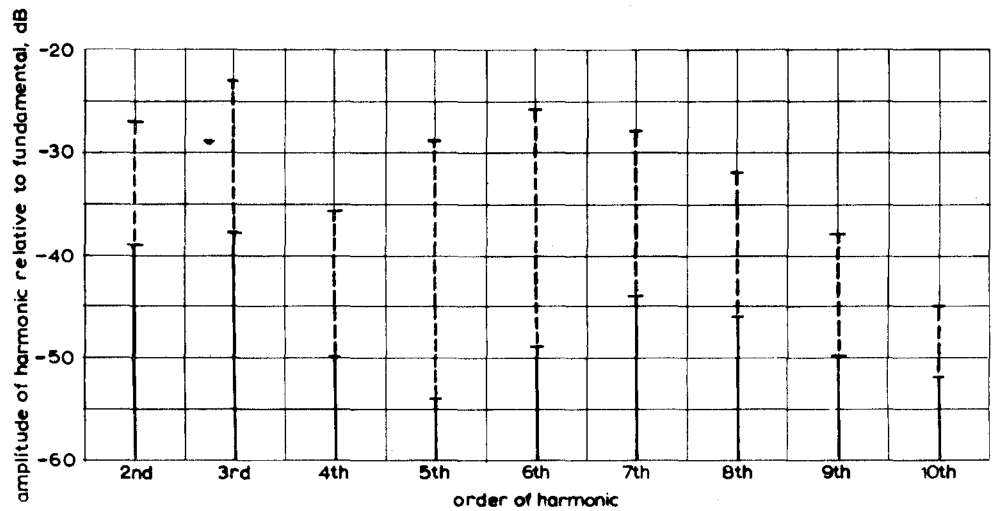

Fig. 17 - Effect of sub-carrier limiting on multipath propagation distortion

——— With adequate sub-carrier limiting
------ With no sub-carrier limiting

adequate sub-carrier limiting and no limiting in the decoder corresponds to a difference in the subjective rating of the interference of about one grade in the scale 'Imperceptible', 'Just perceptible', 'Perceptible', 'Slightly disturbing' and 'Disturbing'.

3.4. Comparison of Monophonic and Pilot-Tone Systems for Susceptibility to Interference

It is not possible to give exact figures for the comparison of performance of a stereophonic with a monophonic system in the presence of the forms of interference considered above. The effects are determined partly by the characteristics of the interference and partly by the design of the receiver. However, in the author's experience, the following represents the average performance attainable, assuming a well-designed receiver; the importance of an adequate design cannot be too heavily stressed.

(i) *Adjacent-channel interference*

In terms of the wanted-to-unwanted signal ratio at the receiver input for 'perceptible' interference, the pilot-tone system requires an additional protection of about 15 dB when the interfering transmission is offset by 100 kc/s but negligible extra protection, not more than 2 to 3 dB, with frequency offsets of 200 kc/s or more.

(ii) *Multipath propagation*

The effect of multipath propagation varies rapidly with the relative delay of the reflected signal. When this delay is comparatively long, corresponding to a path difference greater than about 8 miles (13 km), the subjective assessment of the resulting distortion averages some ½ grade worse, in the subjective scale referred to earlier, with the pilot-tone system. As the path difference is reduced, the difference between the stereophonic and monophonic systems increases from 1 grade at 5 miles (8 km) to 3 grades at 2 miles (3·2 km). In the case of a 2-mile path difference, this does not necessarily mean that distortion will be widespread, since reflected signals with such a short path difference are rarely large enough to cause audible distortion in the monophonic case. The greatly increased sensitivity in stereophony means, however, that the shorter path-differences will become important in a number of cases, but there is insufficient experience to make a practical comparison with monophony.

(iii) *Impulsive interference*

With impulsive interference at a low level the output signal-to-noise ratio is some 20 dB worse with stereophony, as might be expected, since the effect is similar to that of random noise. This corresponds approximately to two grades on the subjective scale. As the impulse level is increased this disparity reduces and, in the practical situation with impulses covering a wide range of amplitudes, the overall subjective difference is about one grade.

All of the comparisons given above apply to stereophonic reception. Compatible reception of the stereophonic transmission on a normal monophonic receiver would be degraded only to a negligible extent.

4. CONCLUSIONS

If a stereophonic v.h.f. sound broadcasting service were instituted in the United Kingdom it would be necessary, in order that the service areas of the transmitters should not be unduly restricted, that stereophonic receivers should be designed to minimize the effects of interference. The requirements for the tuner portion of the receiver are similar in type to those for monophonic receivers although more stringent in degree. The requirements of the decoder, as illustrated in the design described, can be met in a comparatively simple and inexpensive circuit.

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6. APPENDIX

6.1. Effect of Distortion of Difference-Sideband Spectrum

$\frac{\omega_s}{2\pi}$ = subcarrier frequency

$\frac{p}{2\pi}$ = modulation frequency

$\omega_1 = \omega_s - p$

$\omega_2 = \omega_s + p$

θ_1 = phase advance of lower sideband

θ_2 = phase advance of upper sideband

k_1 = factor by which lower sideband amplitude is changed

k_2 = factor by which upper sideband amplitude is changed

If the correct difference-sideband signal is

$$\sin pt \sin \omega_s t$$
$$= \tfrac{1}{2} \cos \omega_1 t - \tfrac{1}{2} \cos \omega_2 t,$$

the effect of the distortion is to change this to

$$\frac{k_1}{2} \cos(\omega_1 t + \theta_1) - \frac{k_2}{2} \cos(\omega_2 t + \theta_2).$$

The output of a square-wave switching demodulator is then

$$\left[\frac{k_1}{2} \cos(\omega_1 t + \theta_1) - \frac{k_2}{2} \cos(\omega_2 t + \theta_2)\right] \left[\tfrac{1}{2} + \frac{2}{\pi} \sin \omega_s t + \frac{2}{3\pi} \sin 3\omega_s t \; \ldots .\right].$$

Considering only the product terms containing audio-frequency components we have

$$\frac{1}{\pi}\left[k_1 \sin \omega_s t \cos(\omega_1 t + \theta_1) - k_2 \sin \omega_s t \cos(\omega_2 t + \theta_2)\right]$$

or, again taking only the audio-frequency terms,

$$\frac{1}{2\pi}\left[k_1 \sin(pt - \theta_1) + k_2 \sin(pt + \theta_2)\right]$$

$$= \frac{1}{2\pi}\left[k_1(\sin pt \cos \theta_1 - \cos pt \sin \theta_1) + k_2(\sin pt \cos \theta_2 + \cos pt \sin \theta_2)\right]$$

$$= \frac{1}{2\pi}\left[\sin pt(k_1 \cos \theta_1 + k_2 \cos \theta_2) - \cos pt(k_1 \sin \theta_1 - k_2 \sin \theta_2)\right] \qquad (1)$$

Let $k_1 \cos \theta_1 + k_2 \cos \theta_2 = a$

and $k_1 \sin \theta_1 - k_2 \sin \theta_2 = b$

Expression (1) then becomes

$$\frac{1}{2\pi} (a \sin pt - b \cos pt).$$

$$= \frac{-1}{2\pi} (a^2 + b^2)^{\frac{1}{2}} \sin(pt + \phi), \text{ where } \phi = \arctan - \frac{b}{a}$$

Thus the demodulated difference signal is advanced in phase by

$$\phi = \arctan - \frac{b}{a} \text{ and changed in amplitude by a factor } \tfrac{1}{2}(a^2 + b^2)^{\frac{1}{2}}.$$

Where only the amplitudes of the sidebands are changed, the expression for the demodulated signal reduces to $\frac{1}{2\pi}$ $[(k_1 + k_2) \sin pt]$. This gives zero phase error and an amplitude change of $\frac{1}{2}(k_1 + k_2)$ in the demodulated difference signal.

Where only the phases of the sidebands are changed, the expression for the demodulated difference signal reduces to $\frac{1}{\pi} \cos \frac{1}{2}(\theta_1 + \theta_2) \sin[pt - \frac{1}{2}(\theta_1 - \theta_2)]$ i.e. a phase error of $\frac{1}{2}(\theta_1 - \theta_2)$ and an amplitude change of $\cos \frac{1}{2}(\theta_1 + \theta_2)$.

6.2. Effect of Phase Shift of Regenerated Sub-Carrier

The output of the difference signal from the demodulator, if the switching sub-carrier is shifted from the correct phase by an angle ψ, will be

$$(\sin pt \sin \omega_s t) \left[\frac{2}{\pi} \sin(\omega_s t + \psi)\right] = \frac{1}{\pi} \sin pt \left[2 \sin \omega_s t \sin(\omega_s t + \psi)\right]$$

Taking only the audio-frequency term we have

$$\frac{1}{\pi} \sin pt \cos \psi$$

6.3. Effect of Phase Shift and Amplitude Change of Demodulated Sum and Difference Signals

Let the sum signal amplitude be changed by factor r_m
and the difference signal amplitude be changed by factor r_s;
also let the sum signal phase error be α_m, and
the difference signal phase error be α_s.

In the case where modulation input of unit amplitude is applied to the A channel alone, it follows from a geometrical construction of the type given in Fig. 2 that the A and B channel outputs of the receiver are

$$A = \tfrac{1}{2}\left(r_m^2 + r_s^2 + 2r_m r_s \cos\left|\alpha_m - \alpha_s\right|\right)^{\frac{1}{2}}$$

$$B = \tfrac{1}{2}\left(r_m^2 + r_s^2 - 2r_m r_s \cos\left|\alpha_m - \alpha_s\right|\right)^{\frac{1}{2}}$$

and the resulting cross-talk level, expressed in decibels, is

$$-20\ \log_{10}\left(\frac{r_m^2 + r_s^2 + 2r_m r_s \cos\left|\alpha_m - \alpha_s\right|}{r_m^2 + r_s^2 - 2r_m r_s \cos\left|\alpha_m - \alpha_s\right|}\right)^{\frac{1}{2}}$$

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